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POWER PROCESSING

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POWER PROCESSING

FRANCISC C. SCHWARZ

NASA Electronics Research Center



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Introduction

The transfer and transformation of electric power from one form as it appears at the terminals of generators to other forms as required by specific loads, as well as the methods that govern the static and dynamic stability of electric systems as a whole, form an engineering discipline described as electric power processing. Load systems that incorporate delicate sensors and complex electronics are imposing more and more restrictive controls on sources of electric energy. Virtually all of these are active loads. The fact that many require a supply of electric power at fixed, although tightly controlled, potentials obscures the existence of possible problems involving dynamic stability that may be caused by often widely and irregularly varying waveforms of a so-called direct current.

The classical power engineer considers himself fortunate if he can solve the often severe problems of nonconventional electric power generation, and sees his task fulfilled if he can provide a given quantity of electric energy as an amorphous mass of kilowatt-hours in any form. The task of transforming this mass into useful electric energy then involves a transformation of voltage or current waves to acceptable forms.

Communication engineers refer to the transformation of voltage waveforms as signal processing, a category that is understood as encompassing most of the associated analytical work as well. It appears appropriate then to refer to the same concept on the power level as power processing.

Power processing in this sense includes three distinct and interrelated functions:

- (1) The transfer or distribution of power.
- (2) The transformation of magnitude and character of time-varying voltage and current waveforms to other such time-varying or nonvarying forms.
- (3) The materialization of the foregoing functions under the constraint of establishing the static and dynamic stability of the system encompassing all the associated electric system components.

A system often has to perform these functions under severe constraints involving reliability, efficiency, physical weight, size, and adverse environmental conditions. Minimization of weight and size requires utilization of the highest attainable frequency of operation within bounds of feasibility. Reliability and efficiency of the critical components are functions of the frequency of operation and, in turn, determine the upper frequency limit.

The electric systems under consideration are often composed of relatively small entities. The individual power demand of its loads is usually comparable to the power-handling capacity of the power-processing equipment and of the generators. A symbolic block diagram of such an electric system is depicted in figure 1. A number of generators supply raw power to the centrally located power-processing subsystem, which may consist of several individual and dissimilar units. The system provides controlled electric power to the individual load subsystems, as shown at the right. This power may be characterized by unipolar or bipolar potentials, continuous or discontinuous voltage waveforms, or any combination thereof. Excess power may be channeled into storage elements of electric energy such as batteries. This flow of power is therefore governed by the constraints imposed by battery characteristics, which include (1) the control of charge rate, (2) the

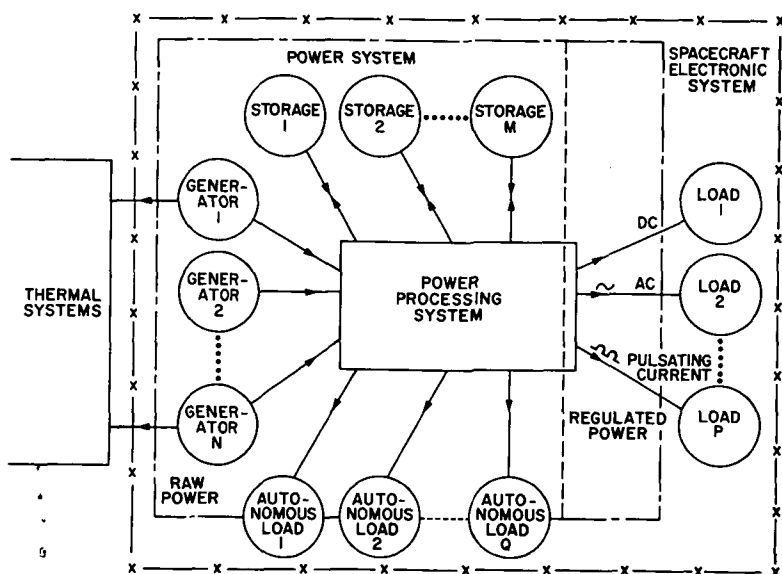


FIGURE 1.—Schematic of electric system.

maintenance of state of maximum charge by repeatedly replenishing charge loss due to "trickle" currents, and (3) the cessation of current supply in presence of excessive battery gas development. Other loads, which appear in the lower part of the schematic diagram, may be referred to as autonomous self-regulatory loads. These loads are powered from sources that undergo wide variations in current or voltage characteristics. Self-regulation is incorporated into the functional mechanism of the load system.

Each of the subsystems forming the electric system indicated in figure 1 has its own characteristics, which govern its static and dynamic stability; thus each must be designed individually. It is also necessary to ascertain overall system stability after interconnection of all the subsystems. The only subsystem connected by channels of power flow to every one of the other subsystems, including the generators, is the power processor. It then becomes mandatory that this (and no other) subsystem contain characteristics that would reject all unacceptable disturbances originating anywhere in the system. Moreover, the subsystem should be able to reconcile conflicting characteristics. Neglect of this requirement may lead to improper system operation or may result in catastrophic failure of subsystems due to imposition of transient, but nonetheless destructive, stresses on component parts.

The designer of the power-processing system should accordingly apply a procedure in which he

- (1) Assembles and recognizes the static and dynamic characteristics of all associated subsystems
- (2) Formulates the needed voltage or current waveform transformations associated with these individual subsystems
- (3) Synthesizes a power-processing subsystem that satisfies the two preceding requirements

The need to comply with these conditions appears to be understood only partially. The designer of power-processing equipment is in many cases a circuit-oriented individual who considers himself a power-conditioning engineer. The term "power conditioning" appears to have been derived from the notion of "air conditioning." Air conditioning has been defined as a process of washing, humidifying and dehumidifying air before it enters a room, hall, building, etc. This analogy suggests that the concept of power conditioning has evolved from the recognition of a regulatory and "purifying" function whereby unregulated power is controlled to remain within required and acceptable limits. In essence, the term "power conditioning" can be understood more in this sense than by extending one step further to encompass wave-

form transformations under constraints of overall systems stability requirements.

The power-conditioning engineer receives, most often, design requirements in the form of input-output characteristics established by systems engineers. These characteristics are, in general, restricted to input-output voltages, power and impedance levels, and other desirable functional requirements. The description of load systems, especially for fixed output potentials, is often fictitiously simplified to that of a resistive load. These loads, though often resistive, are usually not time-invarying, as implied, but consist of active electronic networks that require fixed unipolar potentials while also involving time-varying and irregular current waveforms.

The power-conditioning engineer is aware of the existence of severe problems in his work area, but is often unaware of the fundamental incompatibility between his actual responsibility for the overall electric system and his inadequate approach, which is concerned only with the circuit design aspects. As a rule, he is neither inclined nor trained to assume the responsibility for the stability of inherently nonlinear and time-varying systems of considerable complexity.

Electric Power Transformation

Voltage-waveform transformations are usually implemented by electromechanical or electronic devices. Reliable transformation of waveforms containing significant harmonic components at higher frequencies and extended over a considerable length of time requires the application of solid-state circuit techniques since these circuits are free of limitations imposed by the time constants, vibrations, and wear of mechanical structures.

The classical type of networks used for voltage-waveform transformation is the electronic amplifier. It can be represented as a black box that receives an input, or control signal e_r . At its output terminals, it reproduces this control signal at an enlarged magnitude, at a higher power level, or both. A conventional block diagram of such an amplifier is presented in figure 2. The input-output relationship is simply stated as



FIGURE 2.—Conventional block diagram of electronic amplifier.

$$e_o(t) = Ae_r(t) \quad (1)$$

This relation holds if (1) the supply voltage $e_s(t)$, which is not included in figure 2, is time-invariant and (2) all characteristics of components within the amplifier remain time-invariant. The first of these restrictions reveals that this so-called two-terminal-pair network is in reality a three-terminal-pair network, the third pair serving as the port to admit the source power with voltage $e_s(t)$. When

$$\frac{de_s(t)}{dt} = 0 \quad \text{for any } t \quad (2)$$

then $e_s(t) = E_s = \text{a constant}$, which can be lumped with the assumed constant parameters of the amplifier network. Under this particular condition the conventional (and fictitious) presentation of an amplifier as depicted in figure 2 is justified. If, however, the foregoing constraints are removed, as is the case in any practical network supplied from an actual source, then the representation

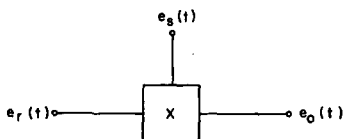


FIGURE 3.—Block diagram of actual electronic amplifier.

of an amplifier is revised to the symbolic form shown in figure 3. The input-output relationship can be now stated as

$$e_o(t) = X[e_r(t); e_s(t); t] \quad (3)$$

That is, the output voltage waveform depends here on the control signal $e_r(t)$, the source voltage $e_s(t)$, and the time-varying character of the operator X expressed by the variable t . The operator X has a time-varying property so that it can compensate for the effects of the time variances of source voltage $e_s(t)$ and of the characteristics of its component parts, so that the simplified relationship of equation (1) can be restored, notwithstanding the fact that $de_s(t)/dt \neq 0$.

In case a unipolar and fixed potential E_o is desired at the output terminals, one can resort to an emitter-follower circuit, supported by appropriate feedback technique, when operating from a dc source with time-varying potential e_s , provided that $e_s > E_o + v_{ce(\min)}$ at any time. Here $v_{ce(\min)}$ indicates the minimum collector-to-emitter potential drop in the respective power transistor. This technique embodies three significant characteristics:

- (1) It is applicable only for dc voltage stabilization.
- (2) It requires that $e_s > E_o + v_{ce(\min)}$
- (3) A fraction, $(1 - E_o)/e_s$, of the power derived from the source of electric energy is dissipated in this process of regulation.

This technique finds widespread application where conditions (1) and (2) prevail and where the power loss (3) can be tolerated. As a matter of fact, it has found such widespread use that it is at times believed to be the essence of "power conditioning." Nothing is farther from the truth, as will be outlined here.

A transformation of the waveform of the source voltage $e_s(t)$ into a multiple A of $e_r(t)$, as stated by equations (1) and (3), can involve any or all of the following operations:

- (1) A change of character of the waveform. For example, one type of unipolar current waveform can be transformed into another type of unipolar current waveform. Such a case is that of a series of unipolar, half-wave sinusoids with programed discretely varying amplitude and space delivered to a radar or laser system from a source with time-varying unipolar potential. This is illustrated in figure 4(a).

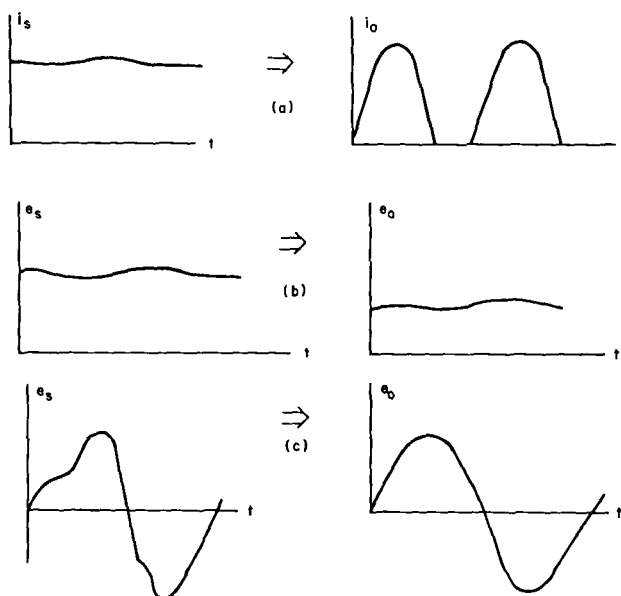


FIGURE 4.—Illustrations of voltage and current waveform transformations (a) in character, (b) in magnitude, and (c) by stabilization.

(2) A process of scaling. For example, one type of unipolar waveform of one magnitude can be transformed to the same type of unipolar waveform of a different magnitude. Figure 4(b) serves as an illustration.

(3) The process of stabilization. For example, a bipolar voltage waveform of quasi-sinusoidal form and time-varying magnitude is transformed to the same type voltage waveform with time-invariant relative and absolute harmonic content, as illustrated in figure 4(c).

Each of these three elements of transformation requires a distinct functional mechanism for its implementation.

Scaling of unipolar waveforms involves usually a dc-to-ac-to-dc inversion and conversion process, which permits the use of transformers for this purpose. Stabilization, on the other hand, requires the application of feedback techniques. Finally, change of character of the voltage waveform results in a different pattern of power flow P_o from the one P_s that would result from connecting the source with potential $e_s(t)$ to a time-invariant, passive, resistive load, as illustrated in figure 5(b). Here reference is made to the waveform transformation indicated in figures 4(a) and

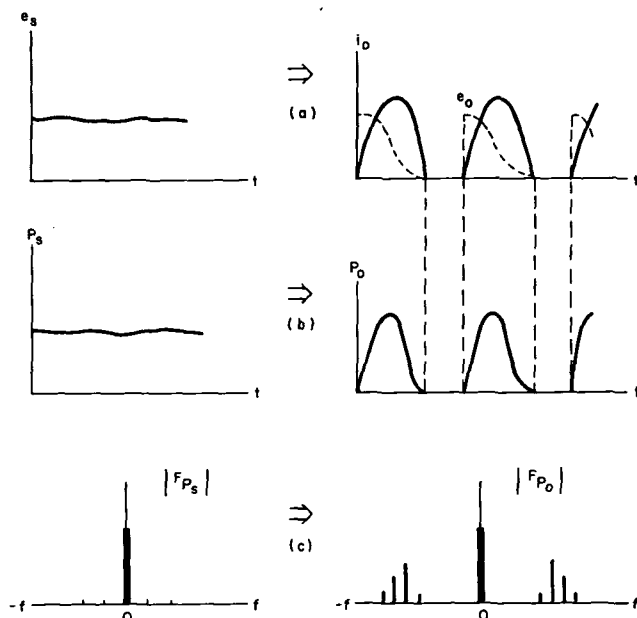


FIGURE 5.—Illustration of (b) power waveforms and (c) frequency spectra under (a) voltage waveform transformation.

5(a). The output voltage waveform $e_o(t)$, as received by a load requiring pulsating currents $i_o(t)$, is indicated by a dotted curve in figure 5(a). The power-frequency spectrums $|F_{P_s}|$ and $|F_{P_o}|$ pertaining to the input and output waveforms, respectively, are depicted in figure 5(c) and indicate the extent of transformation.

The discrepancy between the power waveforms in figure 5(b) reveals the necessity for a continuous exchange of energy between the source and energy-storage components in the power processor and also between these components and the load to bridge the gaps between the two power waveforms. It becomes immediately apparent that a transformation of the character of a waveform may impose a substantial demand for energy-storage components in the power-processing network. This demand is proportional to the extent of transformation of waveform character that can be approximated by the ratio b_{i_s}/b_{i_o} of the normalized coefficients of the lowest harmonic components of the source and output power spectra $|F_{P_s}|$ and $|F_{P_o}|$, respectively, as depicted in figure 5(c). The same demand on energy-storage components is, in general, inversely proportional to the lowest frequency of operation of the power processor.

The reasoning in the preceding paragraph tacitly assumed that the power-waveform transformation should occur in an essentially nondissipative manner. This transformation would otherwise reduce to a simple process involving the dissipation of unwanted power in a time-varying series resistance, such as a controllable semiconductor component. It is noted in passing that even if such a wasteful process could be tolerated, it would constitute a substantial technological problem at power levels in the kilowatt range. In virtually all practical applications, such an approach appears unacceptable from several points of view, such as size, utilization of the source of energy (including its fuel), the bulk of associated equipment (including heat-transfer facilities), and considerations of economy.

NONDISSIPATIVE WAVEFORM TRANSFORMATION

The block diagram of a three-terminal-pair network, as shown in figure 3, can be viewed as that of an amplifier with signal flow from left to right, where $e_r(t)$ is the input signal and $e_o(t)$ is the response. The presence of yet another signal $e_s(t)$ emanating from the source of electric energy is then tacitly ignored. The same diagram can represent a controllable nonlinear active filter that transforms the signal $e_s(t)$ into a desired output signal $e_o(t)$. The latter formulation appears to explain more impressively the actual mechanism of waveform transformation. Such nonlinear active filters incorporate components that will change their transfer characteristics subject to preconceived conditions, which include the effects of a control signal $e_r(t)$. The most common type of these filters is indicated in figure 6(a). Here the nonlinear element is represented as a time-varying resistance, such as transistors in class A operation, and signifies the dissipative character of its power transformation capacity. Figure 6(b) shows the nonlinear element as a controllable on-off switch with, ideally, zero resistance in its closed state and infinite resistance in its open

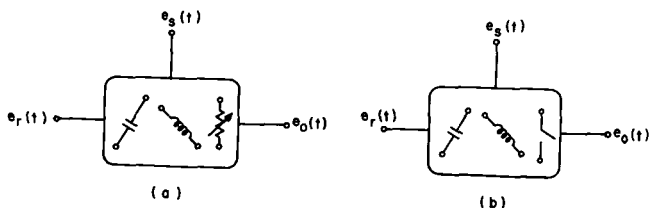


FIGURE 6.—Block diagrams of nonlinear active filters with (a) dissipative controllable elements and (b) nondissipative controllable elements.

state. The waveform transformation is then governed by nominally lossless energy-storage components, such as capacitors and inductors, and the time-varying mode of operation of equally lossless quasi-digital switches. The term "lossless" is defined throughout as "nominally lossless with minimum of residual resistive properties as physically implementable." Such a definition excludes the use of any resistive element or component thereof for any functional purpose in the process of waveform transformation. The further improvement of electric power conversion networks is then not limited by a necessity to preserve the resistive features of certain components for any functional purpose. The basic philosophy of such a nonlinear active filter is discussed with reference to the symbolic diagram of figure 7. A controllable sampling switch CSS samples the source function $e_s(t)$. The mode of sampling is governed by control network H, which acts subject to a control signal $e_r(t)$ and the pulse train $e_s^*(t)$. Filter F removes the undesired harmonic components contained in the pulse train $e_s^*(t)$. If the weight of the k th sample at time t_k is $A_k e_s(t_k)$ and the k th interval between the leading edges of samples is called T_{ok} , then

$$t_k = \sum_{j=1}^k T_{oj} \quad \text{for } T_{om} \neq T_{on} \quad (4)$$

and

$$e_s^*(t) = \sum_{k=1}^{\infty} A_k e_s(t_k) \delta(t - t_k) \quad (5)$$

The delta Dirac function is here introduced as convenient substitute for rectangular pulses, neglecting the effect of harmonic com-

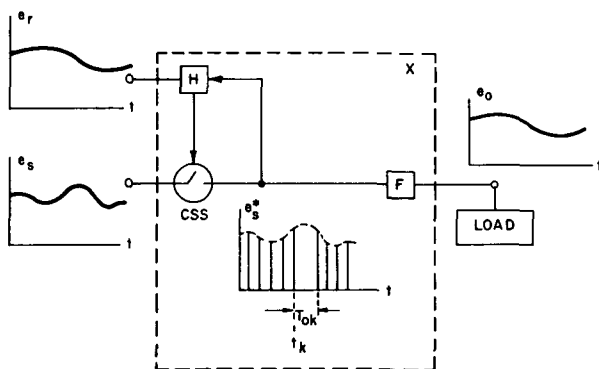


FIGURE 7.—Diagram of nonlinear active filter X consisting of a controlled sampling switch CSS, controller H, and time invariant linear passive filter F terminated in a load.

ponents with frequencies higher than $f_{s(\min)} = 1/T_{ok(\max)} = k_c \omega_c / 2\pi$. The control function $e_r(t)$ is described by

$$e_r(t) = \sum_{n=-\infty}^{\infty} c_{rn} e^{jn\omega_1 t} \quad (6)$$

provided that $e_r(t) = e_r(t + 2\pi m/\omega_1)$, where $m = 0, 1, 2, \dots$; that is, $e_r(t)$ is periodic, with ω_1 as its lowest radial frequency and $\omega_c = n_c \omega_1$ as its highest significant harmonic component. The lowest sampling rate $f_{s(\min)}$ is related to the frequency $f_c = \omega_c / 2\pi$ by the ratio $k_c < 1$, as stated previously. The source function $e_s(t)$ is at this time assumed to be periodic, such that $e_s(t) = e_s(t + m\gamma 2\pi/\omega_1)$ with γ being any positive integer. It will be seen further on that the restrictions imposed on $e_r(t)$ and $e_s(t)$, which require that these functions be periodic, can be removed.

For a number N_β of samples taken from function $e_s(t)$ per period $2\pi/\gamma$, the n th Fourier coefficient $C_{snn}(\alpha)$ of the truncated function $e_{st}^*(t) = e_s^*(t)$ for $2\pi\alpha/\gamma < t < 2\pi(\alpha+1)/\gamma$ is

$$\begin{aligned} C_{snn}(\alpha) &= \frac{\gamma}{2\pi} \int_{2\pi\alpha/\gamma}^{2\pi(\alpha+1)/\gamma} A_k e_s(t_k) \delta(t - t_k) e^{-jn t} dt \\ \text{or} \quad C_{snn}(\alpha) &= \frac{\gamma}{2\pi} \sum_{k=\sum_{\beta=0}^{\alpha} N_\beta}^{\sum_{\beta=0}^{\alpha+1} N_\beta} A_k e_s(t_k) e^{-jn t_k} \end{aligned} \quad (7)$$

$$\alpha = 0, 1, 2, \dots, \gamma - 1; N_o = 1$$

where one lets $\omega_1/\gamma \rightarrow 1$ and normalizes the time accordingly. If, furthermore

$$C_{snn}(\alpha) = C_{snn}(\theta) \quad \text{for} \quad \theta = 0, 1, 2, \dots, \gamma - 1;$$

then C_{snn} is independent of α and can be treated as the Fourier coefficient of a function $e_s^*(t)$ with periodicity $2\pi\alpha/\gamma$ including $2\pi\gamma/\gamma$, although $e_s(t) \neq e_s(t + 2\pi\alpha/\gamma)$. Filter F is designed to eliminate the harmonic components having order numbers beyond n_c . The significance of equation (7) is thereby reduced to the order numbers $n = 0, \pm 1, \pm 2, \dots, \pm n_c$. From equation (6), and after normalization of time, we obtain

$$C_{rnn} = \frac{\gamma}{2\pi} \int_{2\pi\alpha/\gamma}^{2\pi(1+\alpha)/\gamma} e_r(t) e^{-jn \gamma t} dt \quad (8)$$

If we succeed in sampling the function $e_s(t)$ in such a manner that

$$\int_0^{2\pi} e_r(t) e^{-in\gamma t} dt = \sum_{\beta=0}^{\alpha+1} N_{\beta} \sum_{k=\sum_{\beta=0}^{\alpha} N_{\beta}} A_k e_s(t_k) e^{-in t_k} \quad (9)$$

for $-n_c \leq n \leq n_c$, as defined before, then

$$e_o(t) = e_r(t) \quad (10)$$

notwithstanding the fact that

$$e_s(t) \neq e_r(t) \quad (11)$$

and

$$e_s(t) \neq e_o(t) \quad (12)$$

If $\omega_1 \rightarrow 0$ and $N \rightarrow \infty$, one can readily extend the same philosophy to nonperiodic functions. This presentation was chosen for simplicity to clarify the relation between the process of finite sampling and the waveform transformation. Satisfaction of equation (9) depends on the choice of weighting factors A_k corresponding to the finite width of one sample and T_{ok} corresponding to the interval from sample to sample. Proper choice of these parameters is the key to deterministic waveform transformation.

Transformation of a voltage waveform by application of a sampling process appears as the converse of the classical objective associated with these processes, which eventually involves the retrieval of information contained in the original signal or waveform. This applies to uniform and periodic sampling processes widely used in communication and control systems work. The capacity of information retrieval expressed in quantitative terms was formulated by Shannon (ref. 1). It was subsequently shown that application of either a nonuniform periodic or a uniform aperiodic sampling process, or a combination of both would transform the frequency spectrum of the sampled function into another; quantitative relationships between the frequency spectra of Fourier transformable waveforms before and after application of nonuniform or aperiodic sampling have been established as functions of the mode of sampling expressed in time-domain terms. Moreover, it was shown that this process could be applied in a deterministic manner by proper control of the weighing factors A_k or by control of the sampling interval T_{ok} , both having been pre-

viously defined (ref. 2). The various methods of pulse modulation presently used for voltage waveform transformation in the processing of electric power will be considered next.

METHODS OF WAVEFORM TRANSFORMATION

Voltage- and current-waveform transformation with application of dissipative techniques are well covered in the literature on conventional electronic amplifiers. The emitter-follower-type series regulator has already been briefly mentioned; the nature of its control mechanism is discussed in the section on control techniques.

The basic forms of pulse modulation used in power processing networks are illustrated in figure 8. The relationships between the duration T_k of the individual pulses and their associated cycles T_{ok} are indicated in table I. The pulse trains shown in figures 8(b) to 8(d) could emanate from a single-ended "chopper" or "time ratio control" circuit as indicated in figure 9. Here one can recognize the functional elements of the nonlinear active filter previously shown in figure 7. The control box H was retained for simplicity of presentation.

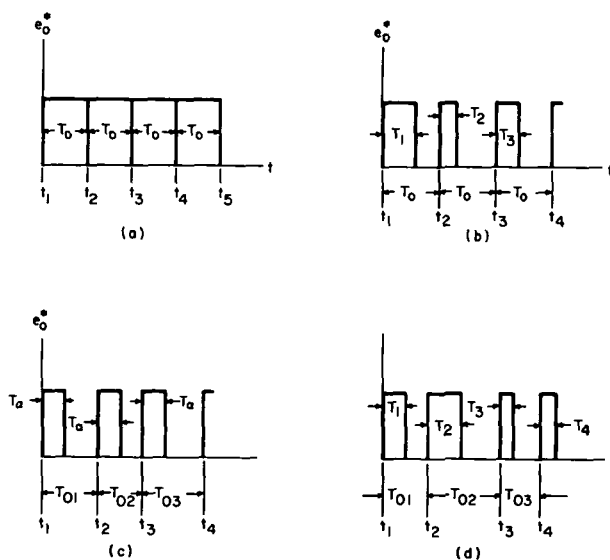


FIGURE 8.—Forms of pulse modulation frequently used in power processing: (a) rectified square wave, (b) pulse width modulated (PWM) rectangular wave, (c) pulse frequency modulated (PFM) rectangular wave, and (d) mixed PWM-PFM rectangular wave.

TABLE I.—Methods of Pulse Modulation for Processing of Power

Mode of operation	Square wave	PWM	PFM PWM-PFM $\left. \begin{matrix} T_m = T_n \\ T_m \neq T_n \end{matrix} \right\} T_{om} \neq T_{on}$
Time intervals ^a T_k, T_{ok}, t_k	$t_k = (k-1)T_o$ $T_k = T_o$	$t_k = (k-1)T_o$ $0 < T_k/T_o < 1$	$t_k = \sum_{j=1}^k T_{oj}$ $0 < T_k/T_{ok} < 1$
$k = 1, 2, 3, \dots$	In processes of inversion, $e_s^*(t_k) = \begin{cases} > 0 & \text{for odd} \\ < 0 & \text{for even} \end{cases} k$		

^a T_k = pulse width, T_{ok} = pulse interval, and t_k = instant of pulse initiation.

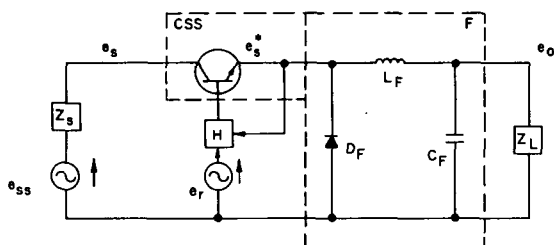


FIGURE 9.—Diagram of typical single-ended pulse modulator for processing of power.

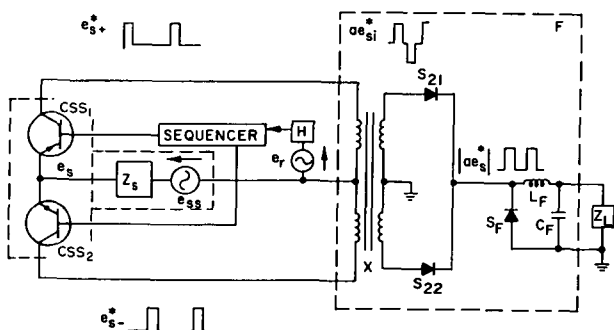


FIGURE 10.—Diagram of power pulse modulator incorporating inversion process.

The same process of pulse modulation is the essence of inverter operation within a converter system of the type shown in figure 10. The controlled sampling switch CSS is here split into two elements CSS_1 and CSS_2 so that a modulated train of pulses e_{s+}^* and e_{s-}^* is fed in successive alternation to the two terminals of a center-tapped transformer. The stepped-up or stepped-down in-

verted voltage waveform ae_{si}^* is subsequently rectified by switches S_{21} and S_{22} and appears in its unipolar form $|ae_s^*|$ at the input terminals of the output filter. If the secondary network of the transformer is reflected to the primary circuit (fig. 10) and the sequence of switching elements CSS_1 , S_{21} , CSS_2 , and S_{22} is appropriately interpreted, then one can obtain the simplified equivalent diagram depicted in figure 7, provided one attributes to filter F the added capability of linear, time-invarying multiplication. It appears that an overwhelming majority of power pulse modulation methods, if not all, can be reduced to the basic functional mechanism presented in and discussed with reference to figure 7. The network illustrated in figure 10 is usually applied to dc-to-dc conversion where the time-varying unipolar source potential e_s is scaled to a predetermined fixed output voltage E_o . That is, the system response

$$\frac{1}{T_{ok}} \int_{t_{k-1}}^{t_k} e_o dt = \frac{aT_k}{T_{ok}} \int_{t_{k-1}}^{t_k} e_s(t_k) dt \quad (13)$$

takes the form

$$\frac{1}{T_{ok}} \int_{t_{k-1}}^{t_k} e_o dt = E_o$$

a constant, provided

$$\frac{T_k}{T_{ok}} = \frac{E_o}{ae_s(t_k)} \quad (14)$$

and

$$de_s/dt = 0 \quad \text{for } 0 < t < T_k \quad (15)$$

Restriction (15) can be removed provided relation (14) is modified to

$$\frac{1}{T_{ok}} = \frac{E_o}{a \int_{t_{k-1}}^{t_k} e_s^* dt} \quad (16)$$

Relation (16) gains significance when one considers that rectangular pulses are a fictitious assumption, if only because in practice the leading and the trailing edges of this type of pulse are neither vertical nor straight. This will be discussed in more detail later in this report.

The desired output voltage E_o and transformer turns ratio a are predetermined constants, and $e_s(t)$ takes the form of an in-

dependent variable. It suffices then to let only one of the time intervals T_k or T_{ok} vary in order that relation (14) or (16) be satisfied, a requirement that can be met by application of either pulse width (PWM) or pulse frequency modulation (PFM). It is evident, and also indicated in table I, that the mixed PWM-PFM process allows two degrees of freedom by varying both intervals T_k and T_{ok} independently. The significance of this added flexibility of the modulation process will become apparent in context with the physical limitations of components that will be discussed later.

Relation (16) will hold in the cases of one or two discretely varying time intervals when a dc-to-dc conversion process under conditions of cyclic stability is considered. The complexity of treatment of transient conditions becomes apparent when it is realized that relation (16) tacitly assumes that the energy storage elements in the filter maintain individually an essentially fixed level of average energy, this being obvious from the fact that these elements are not represented in the expression. This problem is more pronounced in the case of intended time-varying output voltage waveforms.

The power pulse modulator indicated in figure 10 in the form of a parallel inverter can be applied to the secondary generation of unipolar time-varying voltage waveforms, when operated to satisfy relation (9) and its associated conditions. It will reproduce the control signal $e_r(t) > 0$ with magnitude $e_o(t) = k_e e_r(t)$ under the conditions that $e_s(t) \neq e_r(t)$ provided that $e_s(t) > 0$ and $ae_s(t) > k_e e_r(t)$. The restriction that $e_r(t) > 0$ and thus $e_o(t) > 0$ can be removed by appropriately redefining switches S_{21} and S_{22} . If these switches are provided with controlled bidirectional conduction properties, as implemented by bidirectional thyristors, then this system can be used for the amplification of bipolar signals. Its output will reproduce the control signal $e_r(t)$ independent of variations of the source voltage $e_s(t)$; this property is cited as example of an autonomous load as defined in the introduction.

One significant development is that a secondary sine-wave generator having minimum harmonic content other than at the fundamental frequency. The diagram of such a three-phase sine-wave inverter is indicated in figure 11. The windings a, b, and c of an electromagnetic component can be those of a transformer or of an ac motor. Switches S_{ij} are operated so that the individual windings will "see" application of either an alternating square-wave or cyclically averaged sine-wave potential while also providing the appropriate phase relationship of each winding with respect to the other. When one considers one single phase of the three-

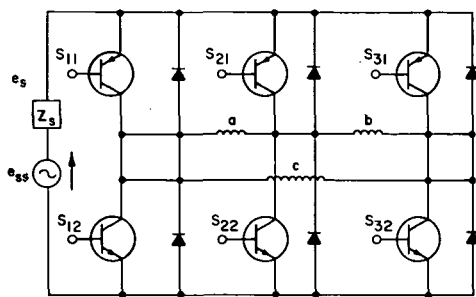


FIGURE 11.—Diagram of three-phase dc-to-ac inverter with motor load.

phase inverter, including the switches associated with that phase, then one recognizes the pulse modulator depicted in figure 10 up to and including the primary transformer winding, except for a difference in implementation in the form of a bridge connection. The voltage-waveform transformation philosophy for the individual phases is, again, identical with the one discussed with reference to figure 7 and expressed with relation (9), although the filtering process may occur inside the motor network, that is, the load.

At this point, the relationship between the methods of power pulse modulation and the objectives of waveform transformation will be discussed.

The dc-to-ac inverter (fig. 11) implements

(1) A change of character of the unipolar waveform $e_s(t)$ to a sinusoidal bipolar waveform that may not be necessarily of fixed amplitude or frequency

(2) A process of scaling of the original magnitude of the unipolar waveform $e_s(t)$ down to a lower amplitude of the resulting ac waveform to allow for the needed variations in the modulation ratio T_k/T_{ok} , unless a transformer with stepup ratio $1:a$ is included as indicated

(3) A process of stabilization by adjusting the average ratio $\langle T_k/T_{ok} \rangle_{\pi f_{ac}}$ so that a constant relationship is maintained between the sinusoidal control signal $e_r(t)$ and the ac rms or average output voltage $e_o(t)$

The dc-to-dc converter (fig. 10) implements

(1) No change of character of the waveform

(2) A process of scaling the waveform, depending on the average ratio T_k/T_{ok} and the transformer ratio $1:a$

(3) A process of stabilization by time-varying adjustment of the ratio T_k/T_{ok} so that relation (16) remains satisfied at any time

A dc-to-dc converter (as referred to above but with the imposed restriction that $T_k = T_{ok} = T_o$; that is, a square-wave inverter converter) implements

- (1) No change of character of the waveform
- (2) A process of scaling the waveform, depending on the transformer ratio 1: a
- (3) No process of voltage stabilization.

The complexity of waveform transformation and of the associated electronic functional mechanisms that depend upon the number and kind of its objectives appears self-evident, even though none of the practical problems involving the physical limitations of components and the associated adverse effects of the severe constraints imposed by physical size and weight, exertion of protective and stabilizing functions, and operation of power processing systems under adverse environmental conditions have been discussed.

An attempt has been made to show that virtually all current solid-state circuit techniques applicable to the processing of electric power are governed by relatively simple pulse modulation philosophies, although their useful analytical treatment appears cumbersome. Several methods of power pulse modulation have been indicated, but even a cursory discussion of the majority of the significant methods in this field is, unfortunately, outside the scope of this report. However, several methods will be introduced to briefly illustrate some important aspects of power processing techniques.

Figure 12 is the diagram of a series inductor inverter converter. The split inductor SI stores energy in its magnetic field until cur-

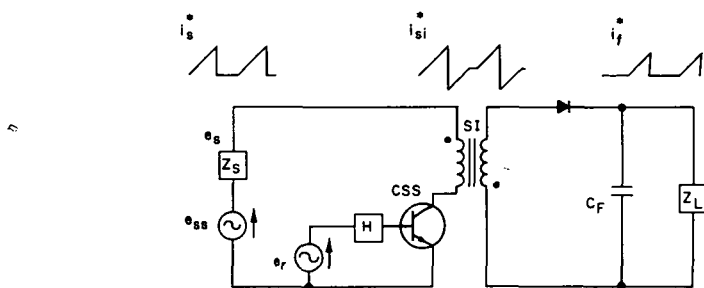


FIGURE 12.—Series inductor inverter converter.

rent i_s^* reaches an intended maximum. After interruption of the current by the controlled sampling switch CSS, the energy stored in the inductor discharges into the filter capacitor C_F and the load Z_L . The inductor SI transfers all of the energy from the source to the load and constitutes a series element in the path of power flow. This current modulation process is analogous to the voltage modulation processes discussed previously.

The diagram of a series capacitor inverter converter in one-half bridge form is indicated in figure 13. The associated waveforms are meant to indicate its mode of operation, as in previous discussions. With this circuit the voltage waveform of the individual pulses is here triangular rather than rectangular, whereas the associated current pulses have a "rectangular" shape. The widths of the pulses are subject to a variation of external conditions, such as the load and source voltage e_s . This represents one method of pulse modulation in which the pulsewidth T_k depends on the time-varying external conditions. The interval T_{ok} must then be adjusted independently so that relation (16) remains satisfied for each cycle of operation (ref. 3). The mode of operation is, therefore, necessarily restricted to mixed PW-PF modulation, illustrated in and discussed with reference to figure 8(d).

Another significant method of voltage-waveform transformation is implemented with the so-called cycloconverter (fig. 14). A three-phase source of ac power operates into an array of controlled sampling switches CSS_{ij} that resembles a three-phase full-wave rectifier system. This array drives a load Z_L preceded by a linear, time-invariant, passive filter F . The frequency of operation of the three-phase voltage source e_s with phase voltages e_{sa} , e_{sb} , and e_{sc} is substantially higher than that of the intended sinusoidal load voltage e_o . Conduction in switches CSS_{ij} is initiated by trigger pulses at times t_k and is terminated by attempted reversal of currents through these switches to zero as polarity of the phase volt-

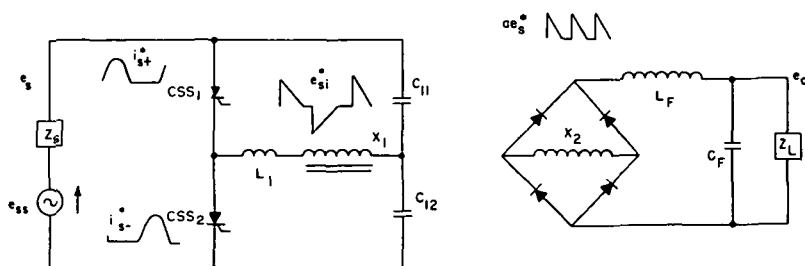


FIGURE 13.—Series capacitor inverter converter.

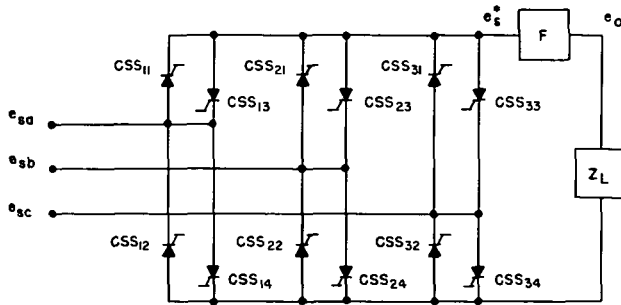


FIGURE 14.—Diagram of three-phase driven single-phase cycloconverter.

ages e_{sm} changes. The control system H selects a certain pair of switches CSS_{ij} associated with phases of appropriate polarity and instantaneous voltage magnitudes to implement a mixed PWM-PFM process to reproduce a sinusoidal control signal $e_r(t)$. The resulting pulse train e_s^* may consist of distinct or immediately adjacent pulses. Relation (16) is then modified accordingly and becomes

$$\frac{ak_s}{k_r} = \frac{\int_{t_{k-1}}^{t_k} e_r dt}{\int_{t_{k-1}}^{t_k} e_s^* dt} \quad (17)$$

where a indicates the transformer ratio $1:a$ and k_s and k_r are constants of proportionality associated with the source and reference voltage waveforms, respectively.

The quality of this transformation is a function of the rate of aperiodic and nonuniform sampling. Relation (17) represents a more generalized form of expressions (14) and (16); it is readily reduced to the latter by reintroduction of the constraints that $de/dt \rightarrow 0$ for $t_{k-1} < t < t_k$ (ref. 2, pp. 34 to 80).

CONTROL TECHNIQUES

The simplest power processor, the dissipative series regulator, is traditionally governed by a type 0 feedback control mechanism (ref. 4). A block diagram of an emitter follower used as a dc series regulator is shown in figure 15. This emitter follower maintains an output voltage $e_o = k_b e_r - v_B - v_{be}$ when $A = 0$, where v_B is the voltage drop of the buffer stage and v_{be} is the voltage drop from base to emitter of transistor Q. The summing operator Σ_2 compares the appropriately scaled reference and output potentials $k_r e_r$ and

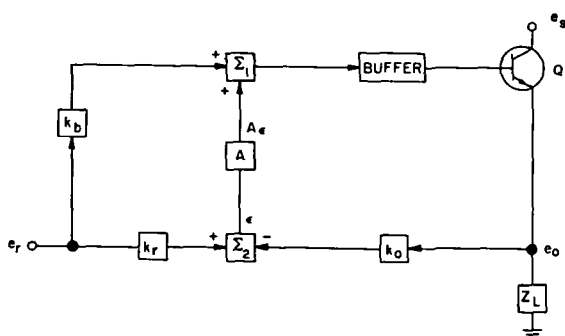


FIGURE 15.—Block diagram of series regulator with type 0 feedback control.

and $k_o e_o$. The output voltage error $\epsilon = k_r e_r - k_o e_o$ is multiplied by the linear factor A and added to the reference drive $k_b e_r$. Deviations of the output voltage from its intended value, which may be caused by variation of voltage v_{be} or other component characteristics, is then reduced by a factor of approximately A provided $A \gg 1$. This three-terminal-pair network then represents a dissipative-type implementation of an amplifier as schematically indicated in figure 3. Its time-dependent transfer function adjusts itself to the varying conditions of the source voltage $e_s(t)$, the load Z_L , and the characteristics of power transistor Q . It is, however, tacitly assumed that the component characteristics of the signal-level control network will not vary because of changes in operating temperature and aging or that the net effect of these variations remains within predetermined limits of tolerance.

Series regulator networks are often used to protect the load from excessive currents, to reject too small or too high values of e_s , as well as reversed polarity of the source. These added functions increase the complexity of this type of network. The dc series regulator is presently widely used, especially as a low-power dc voltage stabilizer within the individual circuits of electronic load systems. Its application, however, is becoming more and more restricted to this type of use because of its relatively low efficiency and its inherent limitation to dc-to-dc conversion with $e_o < e_s$.

Control of nondissipative types of power processors requires conversion of analog signals emanating from the power circuit (e_o) and control waveforms (e_r) to discrete time intervals (T_k , T_{ok}). This process is herein referred to as analog-to-discrete-interval conversion (A/DI). The classical approach to this problem is illustrated in figure 16. The A/DI converter transforms the sum of analog signals $U(t) = k_r e_r + A \epsilon$ to a succession of 1-0 output

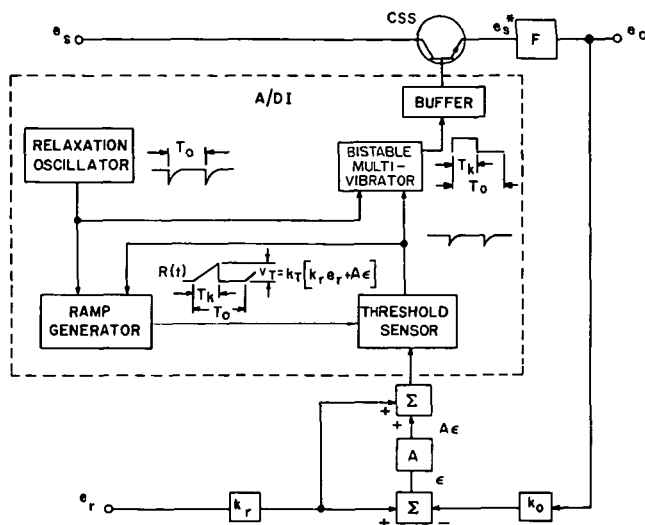


FIGURE 17.—Block diagram of pulse modulator with type 0 feedback control mechanism.

verter. A bistable multivibrator determines the ON or OFF state of sampling switch CSS by a buffer stage. The relaxation oscillator serves as a free-running clock for this pulse-modulated system; it emits a train of trigger signals at periodic intervals to cause a change of state of the bistable multivibrator and initiate the individual power pulses. The same train of trigger signals initiates the emission of individual ramp signals by the ramp generator. The threshold sensor emits a signal whenever the potential of the ramp signal $R(t)$ reaches its sensing voltage v_T , determined by the relationship $v_T = k_T [k_r e_r + A\epsilon]$ where k_T is a proportionality constant. The signal emitted by the threshold detector resets the bistable multivibrator and terminates the power pulse. It also resets the ramp generator. The duration T_k of the ramp signal $R(t)$ emitted by the ramp generator is a function of the detection level v_T for constant slope $R'(t)$. This level is controlled by the feedback mechanism, which thus controls the pulsewidth T_k and the average output voltage level $\frac{1}{T_{ok}} \int_{t_{k-1}}^{t_k} e_o dt$. The functional mechanism of the feedback loop was discussed with figure 16.

Figure 18 illustrates a method that avoids the time delay associated with signal flow through the output filter F. Here the information needed for the pulse-forming process is derived immediately from the switch rather than retrieved from the filter output termi-

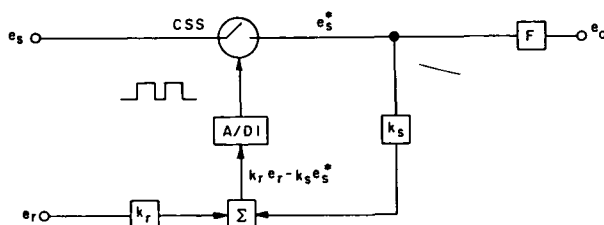


FIGURE 18.—Diagram of self-stabilizing pulse formation in pulse-modulated amplifier.

nals. The advantages gained over the previous method are briefly summarized:

(1) The needless bandwidth limitation of the previously discussed type 0 control system caused by signal flow through the low-pass filter F is largely removed. The resultant increased system speed reduces the former restrictions on e_s , e_r , and e_o and leads to a superior dynamic system behavior.

(2) The problem of dynamic stability is reduced to a fraction of its former significance because the gain A of an outer feedback loop can be substantially reduced, since the signal e_s^* exerts the primary control function. The needed gain for an outer loop feedback amplifier is greatly diminished because of the complete elimination up to the input terminals of the output filter F of the effects of variation of the source waveform e_s , variation of resistive voltage drops inside the converter due to load changes, and variations of the power component characteristics due to changes in operating temperature and aging.

The type of control system in which the k th sample $e_s^*(t_k)$ is compared to the control function $e_r(t)$ at time $t > t_k$ during the entire duration T_{ok} of its cycle and is accordingly formed without any delay has been called self-stabilizing (refs. 5 and 6). It will be seen later that this method includes a process of integration and could be classified as a type 1 control system with zero steady-state and transient error between functions $k_r e_r$ and $k_s e_s^*$ over each closed cycle of operation within the physical limits of valid network function. This system does not, however, correct for errors in e_o due to variations of the resistive voltage drop in the series paths of conduction of filter F . The contribution of this effect to variations in output voltage is a small fraction of the contribution from other causes enumerated before. The resistive component \bar{R}_F of low-pass filters decreases with increasing sampling rate because

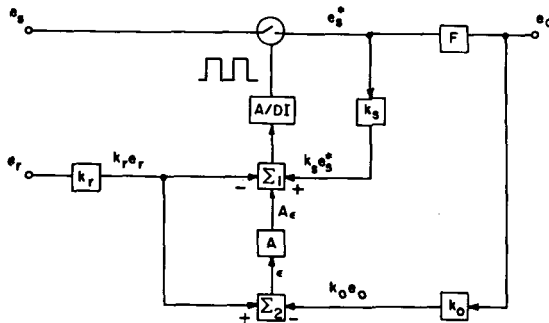


FIGURE 19.—Diagram of self-stabilizing pulse-modulated amplifier with type 1 pulse forming mechanism and type 0 outer feedback correction of residual errors.

small values of series inductance in these filters will satisfy given attenuation requirements. With continuation of the present trends to higher and higher sampling rates to minimize the weight and size of physical apparatus, one could expect eventually to remove the need for correction of this error and thus eliminate one of the primary causes of instability in static power processors. Full-load to no-load variations of output voltage of 2 percent or ± 1 percent due to the resistance to conduction in low-pass power filters for power processors are not uncommon. Reduction of this error to ± 0.1 percent or ± 0.05 percent by gains A of approximately 10 or 20, respectively, represent very modest efforts compared to the substantially higher gains needed for satisfaction of looser tolerances in the feedback control systems discussed with reference to figures 16 and 17. The residual error caused by resistance R_r is reduced by addition of an "outer loop" indicated in figure 19. Satisfaction of relation (17) requires simultaneous solution for the discrete time intervals T_k and T_{ok} . These functions vary from cycle to cycle in discrete steps and are implicitly contained in the upper limits of the definite time integrals of $e_r(t)$ from t_k to $t_{k+1} = t_k + T_{ok}$ and of $e_s^*(t)$ from t_k to $t_k + T_k$, respectively. Solution of this Fredholm equation with two dependent variables and translation of the results into terms of useful electronic signals seems to be a formidable task. There is, however, a relatively simple method for this task. The solution of this problem requires an electronic mechanism that would initiate a cycle of operation at time t_k , observe simultaneously a pulse $e_s^*(t)$ and the control signal $e_r(t)$ from t_k , and emit a signal when relation (17) is satisfied. If relation (17) is rewritten as

$$\int_{t_k}^{t_{k+1}} [ak_s e_s^*(t) - k_r e_r(t)] dt = 0 \quad (18)$$

this will happen exactly at time $t_k + T_{ok} = t_{k+1}$. One form of such an electronic mechanism is indicated in the block diagram shown in figure 20.

The basic control functions required by virtually all the known power pulse modulators can be reduced to the forms indicated in figures 16, 18, and 19; they may, however, require interpretation or expansion for given specific power processing mechanisms. Power processors with switching elements that permit ON-OFF control, such as transistors, require the addition of sequencing logic for the distribution of drive signals to the individual switches in a required order. Mechanisms using switching elements that permit only ON control, such as thyristors, require protective logic to prevent initiation of current flow in any of the switches under any conditions, as long as a short-circuit path could be formed because of the failure of another switch to open before the current flow begins. Acceptance and reproduction of bipolar control signals $e_r(t)$ require the introduction of appropriate means of sensing, interpreting, and directing power flow, as required for dc-to-ac inverters, ac-to-ac converters, and general-purpose amplifiers. Detection of unacceptable conditions in the supply line or the load or internally generated conditions, such as transient or lasting overvoltages or undervoltages and excessive currents, which would endanger critical components, requires electronic sensing and interpretation mechanisms that will direct the control mechanism to protective modes of operation including complete cessation of its functions. These protective systems may include control over redundant sections of power flow or control logic.

Even a cursory outline of these techniques appears beyond the scope of this report. Their existence is mentioned to indicate the kind and level of effort needed in operational electronics for control of power processors. One recognizes the application of basic

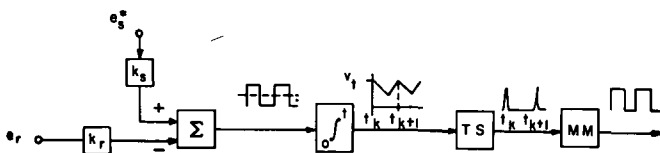


FIGURE 20.—Block diagram of analog signal to discrete time interval converter (ASDIC).

techniques of electronic circuits used in analog and digital signal processing. The need for undelayed response to the continued progress of each cycle of the power system and the issuance of ad hoc solutions on the duration of these cycles with accuracies within small fractions of $1\ \mu\text{sec}$ imposes severe constraints on circuit design. These constraints are further compounded by the fact that there is no room for even a temporary malfunction of the protection system during any of the individual cycles of operation because such a malfunction could lead to the catastrophic failure of the power processor.

Other factors that contribute to the difficulties of design are the absence of an external regulated source of electric energy to power the control system; the proximity of the power circuits of the pulse modulator to the associated sizable and abruptly changing electromagnetic fields; and the disturbance of the prime source of electric energy by the pulsating mode of operation, often compounded by adverse environmental conditions. Further difficulties are introduced by floating reference nodes of individual control system parts that are associated with corresponding sections of the power system and the need for unerring communication between these system parts that "ride" on different continuously or abruptly changing power waveforms. Positive signal transfer, unaffected by relative reference node voltage variations, is needed under transient and static conditions of the system's cyclic operation. This type of operational requirement usually implies direct coupling and imposes the need for application of more sophisticated techniques, in contrast to capacitive coupling that can be freely applied in standard signal-level electronics.

Philosophy and techniques of operation of power processors have been outlined in the preceding two sections. The rate of development of both has been continually dominated by problems in the areas of networks and components. These problem areas, and their effect on the efforts in the field of power processing, are presented in the following section.

Problem Areas

CHARACTERIZATION OF PROBLEMS

The feasibility and reliability of static power processors depend largely on the availability of semiconductor power components with characteristics that can satisfy the requirements of their power circuits. Instant or delayed catastrophic failure of these components can occur when component ratings, such as maximum voltage, current, or allowable power dissipation, are exceeded for even relatively short times. Thus, these semiconductor switching components must be selected to endure the largest expected stresses due to applied potentials, currents, and heat dissipation. Estimates on the magnitudes of these stresses are based on intended steady-state operation with added allowances for irregular source and load conditions and unintended cyclically recurring and transient voltage and current excursions including spikes.

Although the needed characteristics of switching elements based on the required power handling capability of the network can be readily calculated and established, no analytic methods have been developed to calculate the other stresses that are caused by dynamic system behavior and by the physical properties and geometrical configuration of component materials. Attempts to treat these areas quasi-analytically appear restricted to application of iterative computerized numerical analysis, referred to as network simulation or computer-aided circuit design. These approaches will permit, at best, the acceleration of trial-and-error processes but are burdened with the inherent limitations of incomplete device modeling. As a result, the designer finds himself in the position of having to derate the switching components substantially to allow for contingencies whose existence is known but whose magnitudes are rarely known. Derating is roughly proportional to the degree of reliability of operation and can reduce the power handling capacity of a power transistor by an order of magnitude when ultrareliability is required, as with power conversion equipment on a manned spacecraft.

Even then the designer does not know what margin of safety, if any, is left or whether his design is inherently unreliable. He

needs information on the interdependence of the nonlinear characteristics of components interacting in circuits that function in nonlinear modes, such as those that implement pulse modulation processes. His task is compounded by the fact that satisfactory analysis of these networks would still be cumbersome, even if this interdependence of component characteristics were better understood—there are no established and practical methods of analysis.

SWITCH-INDUCTOR-TRANSFORMER INTERACTION

One of the classic problems encountered in power pulse modulation is that of interruption of current in an inductor. In fact, flow of current in an inductor cannot be abruptly terminated, and the attempt to do so will lead to a breakdown of other components unless another path of current flow is provided.

Virtually all pulse-modulated power processors incorporate inductive elements in their circuits that can be either intended or parasitic, unlike their signal-level electronic counterparts. When the current is interrupted in the course of transferring the stored magnetic energy to either the load or capacitive elements, a fraction of this energy is dissipated in the parasitic resistive elements of the circuit. The distribution of the resulting conversion of the magnetic energy to heat in the circuit is proportional to the relative magnitude of the resistive elements within this circuit. The largest of these resistive elements is invariably the switch in the process of opening. This discharge of magnetic energy manifests itself in the form of voltage and current spikes that appear at the termination of individual cycles of power pulse modulators. Many designers of power processing equipment are well aware of these phenomena, which are usually but not always due to parasitic effects associated with power components and networks.

The parallel inverter, illustrated in figure 10, incorporates at least two functions that can be cited as examples. The currents flowing alternately through the primary inverter circuits are terminated by opening of switches CSS_i . These switches, in effect, interrupt the current flowing in filter inductor L_F . The problem, then, is the transfer of the current from switch S_{2i} to S_F concurrent with the interruption of the same current by the controlled sampling switch CSS_i . The diagram of the network under consideration should be modified to replace the simple switch symbols by their more representative equivalent circuits, and to include the secondary characteristics of the inductor to permit a detailed transient analysis of this process under the given "large signal" conditions. To understand the processes involved and to design an

optimum system, the nonlinear characteristics of the elements involved during their respective transitions through delay, rise, storage, and fall times require careful modeling and close understanding and correlation of the nature of these modeling processes with each other during all phases. This study should clarify the phenomena associated with these transitions, including the waveforms of voltage and currents at the component terminals, and the time and spatial distribution of these potentials, currents, and resulting heat dissipation within the switches. This information would provide a significant contribution to the understanding of stress mechanisms and permit judicious application of design criteria. No successful analysis of this type has been reported.

The real characteristics of the power transformer X of the parallel inverter have not been considered up to this point. This apparently simple device assumes considerable complexity during abrupt current flow transitions within time intervals on the order of microseconds (ref. 7). The simple transformer symbol in figure 10 is replaced by an equivalent circuit to fit the conditions during abrupt transitions and other conditions. These equivalent circuits include lumped parameters to approximate the effects of distributed and time-varying electromagnetic fields that are due to interwinding and intrawinding capacitances, leakage inductance, apparent power losses in ferromagnetic core materials, skin effects, and induced eddy currents in conductors, to name some significant phenomena. The analysis of even the simplest inverter network containing all the significant parameters of semiconductor and electromagnetic wirewound components is very complex. The interaction between the transistor switch in the primary circuit of a parallel inverter and the magnetic energies stored in the leakage inductance and the airgap of the magnetic core of a power transformer is briefly considered as an example for such an analysis.

The forcible interruption of current flow at the termination of each pulse is initiated by application of a turnoff signal to the base terminal of transistor CSS₁. The actual opening process is initiated after the transistor's storage time has elapsed. The magnetic energy stored in the leakage inductance of the primary circuit and in the airgap of the transformer is unable to return to the network inconspicuously, as it does under conditions of sinusoidal excitation. The stored magnetic energy will attempt to adjust the potentials of transformer terminals so that this energy can be transferred to the associated circuits. This process necessarily leads to dissipative conversion of electromagnetic energy to heat at times when this energy is not transferred to other energy

storage elements for potential reuse. The return of the magnetic energy to the primary inverter circuit manifests itself by the occurrence of voltage and current spikes at the trailing edges of each individual pulse appearing across the terminals of switches CSS_i (fig. 21). These spikes interact with the opening process of the switch and modify substantially the switch component characteristics that otherwise appear within resistance-capacitance (RC) networks in which the characteristics are usually established. It is expected that a reduction of these spikes will alleviate the stresses on the switching component.

One way to reduce these sudden waveform excursions is to reduce the amount of returnable energy stored in the magnetic fields of the transformer. Such a reduction could be accomplished by the complete elimination of the airgap of the transformer core and the use of ferromagnetic core materials that are characterized by "square" and narrow BH loops. Use of such cores eliminates the airgap problem and reduces the leakage inductance of transformers considerably. It introduces, however, the problem of saturation currents due to the absence of the airgap that otherwise centers the BH loop around the origin of the BH coordinates. This problem is compounded by the delay interval T_d between the instant of detection of transformer core saturation at time T_f and the actual completion of the opening process of switch CSS_i at time T_k . The typical magnetizing current i_m of the transformer normalized with respect to its nonsaturated average magnitude i_{mn} is depicted in figure 22. Operation of inverters involving these current waveforms was first introduced by Royer (ref. 8) and later modified by Jensen (ref. 9).

The interaction between the parasitic effects in high-frequency

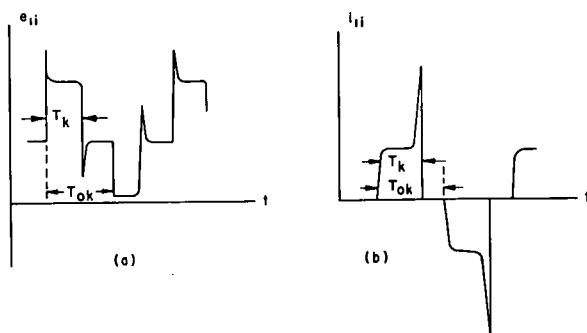


FIGURE 21.—Waveforms at switches CSS_i of parallel inverters: (a) voltage and (b) current.

power transformers and the physical shortcomings of semiconductor switches constitutes one of the most serious problems in present power processing systems. The delay T_d in opening switch CSS_i causes an increase of the magnetizing current toward $i_m(T_k)$, a multiple of the load current i_L even though the ratio of the "non-saturated" magnetizing current to the intended load current $i_{mn}/i_L < 1$ by virtue of required transformer efficiency. The increased primary current $i_{1i}(T_k) = i_m(T_k) + i_L$ then increases from its "normal" value $i_{1in} = i_L$ to a multiple $k_m \gg 1$ such that $i_{1i}(T_k) = k_m i_{1in}$ as indicated in figure 21. The energy stored in the leakage inductance at time T_k is then increased by the factor k_m^2 from its former value when $i_{1in} = i_L$ provided sufficient base drive is applied. The switch opening process is thus further complicated.

This problem has been attacked in two ways. The first attempt was to devise and implement a pulse modulation philosophy that compensates for the modulation error caused by the turnoff delay T_d by adjustment of the aperiodic cycle duration $T_{om} \neq T_{on}$; the second was to center the BH loop of saturable core transformers around the coordinate origin. The required type of modulation is provided by control mechanisms discussed with reference to figures 18 to 20. The significance of the last statement is explained by reference to the illustration of a flattop pulse in figure 23.

The discrepancy between the preprogrammed (shaded) rectangu-

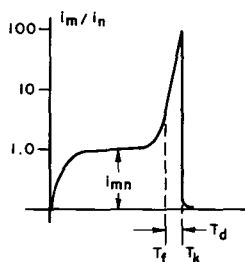


FIGURE 22.—Waveform of normalized magnetizing current i_m/i_{mn} in free-running parallel inverter with saturable core transformer.

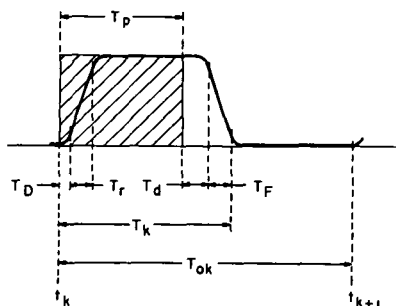


FIGURE 23.—Preprogrammed and actual single flattop power pulse.

lar pulse with duration T_p and the actual pulse as "seen" by the control network is self-explanatory. The functional error that this discrepancy introduces into the control mechanism is removed by appropriate adjustment of the cycle interval T_{ok} to the actual time integral of the k th pulse as expressed by relations (14), (16), and (17) (ref. 2, chs. II and III). This type of control employs mixed PWM-PFM control because only a subsequent adjustment of interval T_{ok} can satisfy relation (17) if there is no way to control the pulse area integral accurately, and PWM with a fixed period for intervals $T_o = T_{ok}$ lacks the required degree of freedom. For these reasons use of mixed PWM-PFM eliminates the need to develop transistors with shorter storage times T_d .

The problem of transformer saturation during switch opening as illustrated in figure 22 has satisfactory solutions (ref. 10). Oscilloscope traces of the current waveforms in the primary circuits of a free-running pulse-modulated parallel inverter before and after application of a new technique are presented in figure 24.

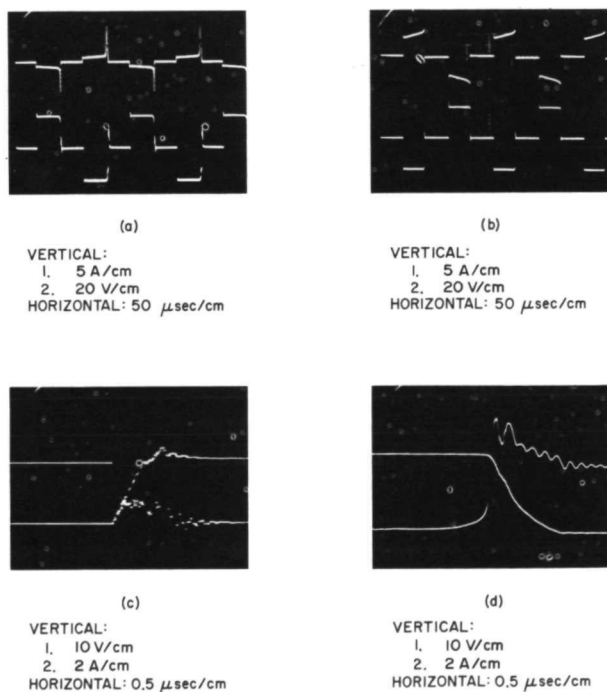


FIGURE 24.—Current and voltage waveforms in primary circuits of free-running inverters. (a) Drive signal removed after initiation of transformer core saturation. (b) Transformer with centered BH loop. (c) Leading edge of current waveform. (d) Trailing edge of current waveform.

The significance of this technique consists in the fact that the current spikes illustrated in figure 24(a) are not suppressed but prevented. The heating effects within the power transistors that are due to this phenomenon are completely eliminated, a significant fact because the localized stresses due to high, fast changing current densities within the transistor constitute a considerable reliability hazard.

As a consequence of this improvement, it can be expected that derating of power transistors could be substantially reduced. The load current carrying capacity of the available switching elements could be increased accordingly to a multiple of its present magnitude. This reduction of the need for derating of switching components should permit design of parallel inverters with substantially higher power capacities than is presently possible.

The series capacitor inverter indicated in figure 13 eliminates unfavorable effects of interaction between transformer and switching elements to a large extent. The current waveform of the individual pulses is that of half sinusoids, which are terminated by turnoff of the silicon-controlled rectifier when the current falls to zero. The current waveform within such inverters is indicated in figure 25. Energy stored in the magnetic fields associated with the pulse currents is transferred to the series capacitor before opening switch CSS_i because of cessation of current flow. Using this type of operation, the switches can be closed and opened with essentially no current flow and freedom from the harmful effects of discharge of magnetic energies during the switch opening process. The series capacitor inverter requires careful design of the electronic control mechanism and an understanding of the physical limitations of the controlled rectifiers because inadequacy of design could lead to immediate failure; in comparison, the parallel inverter may suffer from an initial reduction of efficiency and eventual failure under similar condi-

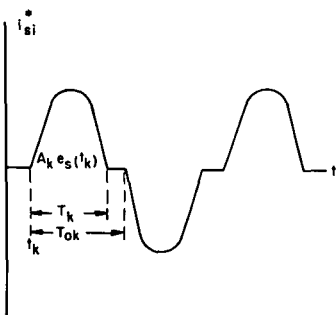


FIGURE 25.—Wave of current i_{si}^* with resonant turnoff in series capacitor inverter.

tions. The inadequacies of series inverters were caused by the necessity to overdesign transformers and inductors substantially so that the associated converter systems could cover a wide range of loading. Recent advances removed this difficulty and led to successful implementation of a 2-kW converter (ref. 11).

ELECTRONIC SWITCHING ELEMENTS

The displacement of rotating inverters and converters by static electric power inversion and conversion equipment was initiated with the advent of electron tubes during the first decades of this century. Moderate efficiency and reliability restricted the use of static inverters to specialized applications, such as induction heating, that could tolerate the associated elevated cost. Frequency of operation was limited to approximately 1 kHz because of the relatively long deionization time in mercury vapor electron tubes. Recent advances in electron-tube technology and the use of gases such as xenon with considerably shortened deionization times permit operation at frequencies of several kilohertz. However, the relatively large forward voltage drop in these tubes, the needed filament heating power, and the fragility of mechanical structures, including that of the filament itself, severely restrict the use of these tubes. The power semiconductor components have continued to displace both rotating and static electron-tube power conversion equipment because of the relatively high efficiency and reliability of these components.

Essentially, the power transistor differs from its signal-level counterpart by the size it requires to carry relatively large currents. The associated larger geometries inherently offset improvements in transistor characteristics, such as fast delay, rise, storage, and fall times, desirable for more accurate pulse control at higher frequencies of operation; better compatibility with transformers and inductors; and minimization of energy dissipation at the fringes of pulses. The relatively thin, fragile base remains a matter of concern because it is believed to be the site of most power transistor failures under conditions of excessive stresses. These stresses lead to localized excessive potential differences or overheating due to temporary concentration of currents.

One way to provide ample current carrying capacities and yet preserve the favorable characteristics of smaller transistors is to put a number of smaller transistors in parallel and enclose them in a common package. The obvious problem associated with this technique is that of load sharing. Conscientious selection and matching of individual transistors tends to reduce this problem.

This solution, however, leaves doubts about the preservation of the match of characteristics throughout the process of aging and introduces an added reliability hazard. The characteristics and problems of transistors are well covered in the pertinent literature. The purpose of this discussion, therefore, is to emphasize the need for useful information on the characteristics of power transistors to the designer of power processing equipment.

Information contained in the specifications of power transistors usually describes this device in terms of characteristics needed for signal-level operation in the linear region of operation. Information of vital significance about component characteristics in the intended ON-OFF switching modes of operation such as delay, rise, storage, and fall times and their dependence on operating conditions is scant, if provided at all. The data that characterize these components are often obtained under unrealistic test conditions, such as purely resistive circuits or, at best, RC circuits. The information contained in the specifications does provide a general orientation on the applicability of the device under consideration. The power transistor, however, is almost exclusively used in networks that include inductors and transformers with significant inductive characteristics. The designer circumvents the difficulty that results from a lack of knowledge of component behavior by derating these switching devices considerably. But he does not know how effective this derating process actually is and has to rely on his experience, intuition, and the outcome of extensive testing of power conversion systems as closed entities.

A better understanding of the interdependence between characteristics of semiconductors and electromagnetic devices such as inductors and transformers seems necessary to provide guidance for the improvement of power semiconductor components capable of functioning in a manner that will fulfill the requirements of solid-state power processing apparatus. Continued efforts for reduction of turnoff times of power transistors below 1 μ sec should be viewed critically because the speed of the turnoff process may be limited by the time constants of the associated electromagnetic components rather than by the electronic switch. These time constants can be considerably larger than the achievable transistor turnoff times and would curtail the usefulness of the fruits of such tedious and expensive efforts.

The silicon-controlled rectifier tolerates relatively irregular operating conditions, such as considerable transient excess stress due to applied potentials or carried currents, and is the most important switching element for high-power static conversion equip-

ment. Its availability with substantial ratings such as breakover voltages on the order of kilovolts and current carrying capacities of several hundred amperes in single units seems to designate this p-n-p-n-type switching element as the workhorse of power processing equipment. This switching element indeed finds application in virtually all power conversion systems that require a capacity greater than $\frac{1}{2}$ to 1 kW for single units. Its application is, however, largely restricted to systems that are being operated from ac sources.

The overriding reason for this restriction is the lack of control over this switch after current flow has been initiated. Alternating-current operation would necessarily lead to the reversal of current flow and thus open the switch. In a dc-powered system this is, of course, not true. A number of techniques have been developed to implement turnoff mechanisms for this device. Turnoff circuits that are externally associated with the controlled rectifiers and built-in mechanisms that permit gate turnoff by removal of the carriers from the junction below the required minimum concentration for conduction are commonly used for this purpose. This type of controlled rectifier operation has found wide application in terrestrial and marine technology where incidental failure of one turnoff process can be remedied by replacement of a fuse. Designers, however, are reluctant to use this switch in applications where continued and uninterrupted operation is vital, such as in aircraft and spacecraft. An exception is the series inverter that positions a series capacitor between the dc source power line and its reference node at any time. However, more work is needed in this area before these systems could become operational.

The controlled rectifier restricts the frequency of operation of power processing equipment to approximately 10 kHz because it needs a recovery time of approximately 10 μ sec. At 10 kHz, this results in a maximum duty cycle T_k/T_{ok} of 0.8. Reduction of the recovery or turnoff time will permit an increase of operating frequency. It should be possible to reduce this turnoff time to the order of 1 μ sec by construction of interdigitated gates and by application of ion implantation techniques. This should permit operating frequencies on the order of 100 kHz and reduction of the physical weight and size of multikilowatt dc-to-dc converters to less than 1 lb/kW while maintaining efficiencies well in excess of 90 percent.

It can be said that the transistor constitutes the presently preferred switching element for power converters with capacities up

to several hundred watts per single unit. Its application is restricted to relatively low input voltage consistent with the relatively low power application. Typical voltages are 28, 36, and, at times, 56 or 80 V. The silicon-controlled rectifier can handle power levels up to approximately 100 kV-A when applied in pairs to single converter units. Its application to dc-driven systems that require a high degree of reliability is subject to successful demonstration of newer techniques. Experimentally acquired data on the efficiency of this type of converter (ref. 11) are indicated in figure 26.

TRANSFORMERS AND INDUCTORS

Transformers and inductors are the key elements in power processing systems. The transformer has the ability to scale potentials and to provide reference node isolation. The inductor is a necessary requisite for all electronic power mechanisms that operate in a pulsating mode. It is the component that will bridge the abrupt and time-varying discontinuity of potentials of adjoining components inside the conversion mechanism and provide the link for power transfer under these conditions.

The technology of transformers and inductors, as currently known, dates back to the decades before the turn of the century.

SERIES CAPACITOR DC TRANSFORMER EFFICIENCY AS FUNCTIONS OF OUTPUT POWER WITH INPUT VOLTAGE AND INPUT CURRENT AS PARAMETERS

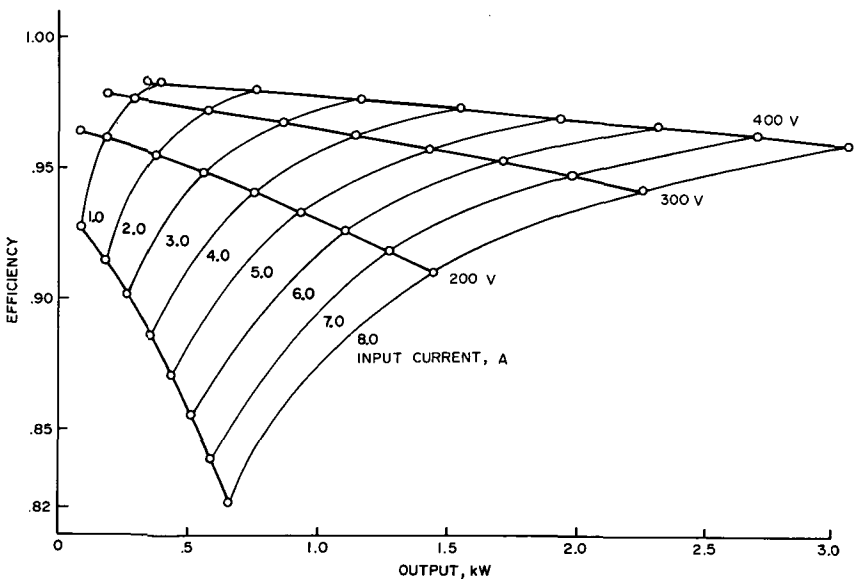


FIGURE 26.—Series capacitor dc transformer efficiency as functions of output power with input voltage and input current as parameters.

Their design and construction have remained almost unaltered except for improvement of materials. The currently applied archaic design methods consist of tedious trial-and-error procedures in which the designer selects magnetic cores by experience and intuition, goes through a number of paper designs in succession, compares them with each other in reference to physical weight, size, efficiency, and cost, and finally selects the one with the most favorable characteristics. Mechanization of these trial-and-error methods by use of digital computers provides some relief. However, the development of comprehensive modularized programs is needed so that the multitude of parameters associated with materials properties, geometric core configurations, and applied waveforms, beyond reported attempts in this field, can be included (refs. 12 and 13). The designer can choose among a relatively large variety of magnetic materials with various performance characteristics available in many geometrical configurations (refs. 14 and 15). He can also choose among various types of core construction, such as a sintered ferrite core or a magnetic core formed from wound thin tape or a stack of laminations, again with a choice of various thicknesses of tape or lamination (ref. 16). The area of standard power transformers for 60-Hz, and to some extent for 400-Hz, operation under conditions of sinusoidal excitation has been relatively well covered over many years. The designer of the transformers and inductors for power processing equipment that usually operates at substantially higher internal frequencies finds himself faced with the task of designing components solely from his knowledge of electromagnetics, including ferromagnetics, for conditions of operation, including voltage waveforms and frequencies, not previously encountered or described in textbooks. Moreover, important material properties like the apparent core loss in magnetic material under conditions of nonsinusoidal excitation are not recorded. Nor, except for very isolated cases (ref. 17), is there a known method to apply the information on core losses under known conditions, such as sinusoidal excitation, to the nonorthodox waveforms that are encountered in power processing systems. This lack of information about how to cope with problems like the magnetic core loss effects due to rectangular voltage waveforms with varying duty cycles having triangular or quasi-sinusoidal waveforms, as encountered in certain types of inverters, forces the designer to attempt to substitute heuristic assumptions for the lack of needed knowledge.

The transformer designer is faced with a similar problem if he tries to evaluate the significance of parasitic effects in wire-

wound magnetic components in detail. The significance of these effects and their interaction with the nonlinear characteristics of semiconductor components have been discussed. An impressive multitude of publications exists in the area of transformer design. A closer examination of this literature reveals that they are concerned with the low- or high-frequency transformer under conditions of sinusoidal excitation or with the wideband or pulse transformer used for specialized application, but not with the power transformer in abruptly switching networks (refs. 18, 19, and 20). It is not unusual for a transformer designer to attempt to place his transformer into a circuit environment that is known to him by experience rather than to venture into an area of innovation. This appears to be true even if that experience conveys to the designer more of a familiar environment than actual knowledge. Besides the lack of essential knowledge, the quest for lower weights and higher efficiencies, and thus higher frequencies of operation, counteracts the possibility of using reflective design procedures and constitutes one of the significant causes of difficulty in this area.

Another problem area is that of transfer of heat generated within the electromagnetic devices. The same amount of heat is generated within the smaller geometrical volume of a transformer when its physical size is reduced at higher frequencies of operation and its efficiency remains unaltered. One of the possible solutions to this problem appears to be a further improvement of the efficiencies of transformers by more extensive application of high-efficiency magnetic materials with narrow BH loops. This in turn requires a better understanding of the involved phenomena and material properties, reinforced by deterministic methods of analysis and synthesis of design.

CAPACITORS

The reliability and stability of dc filter capacitors cause concern in the area of power processing. Electrolytic capacitors packed to high energy densities employ relatively thin electrodes, which constitute one of the causes of insufficient reliability. The successive interchanges of metal between the electrodes and the electrolyte are not necessarily uniformly distributed over the area of interaction. This can lead to the appearance of holes in the electrodes that will eventually enlarge because of the associated field effects. Furthermore, these capacitors suffer from a lack of stability of their capacitance value as a function of temperature. Variation of capacitance in dc filters affects the attenuation capa-

bility of these filters, which produces a variation of ripple voltages. Appropriate overdesign can be applied to limit these effects. However, the effects of varying capacitance on the stability of feedback systems such as power processors are cumbersome to deal with and often require the application of tedious compensation techniques that may vary from unit to unit and with age. The use of all kinds of electrolytic capacitors, including those that contain tantalum materials, has been excluded from certain areas of technology such as manned space flight because their shortcomings could have critical effects. However, even in these areas the lack of other capacitors with comparable energy density forces engineers to use these components and to accept the problems associated with their use, one of which is the adverse effects on reliability and stability of the power processors. Thin-film metal oxide capacitors with energy densities that equal or perhaps exceed those of electrolytic capacitors are expected to improve this situation. Excellent temperature stability and high energy density have been demonstrated by these capacitors. Their uses, however, are only now being explored. (See fig. 27.)

The reason that conventional metal film capacitors have not

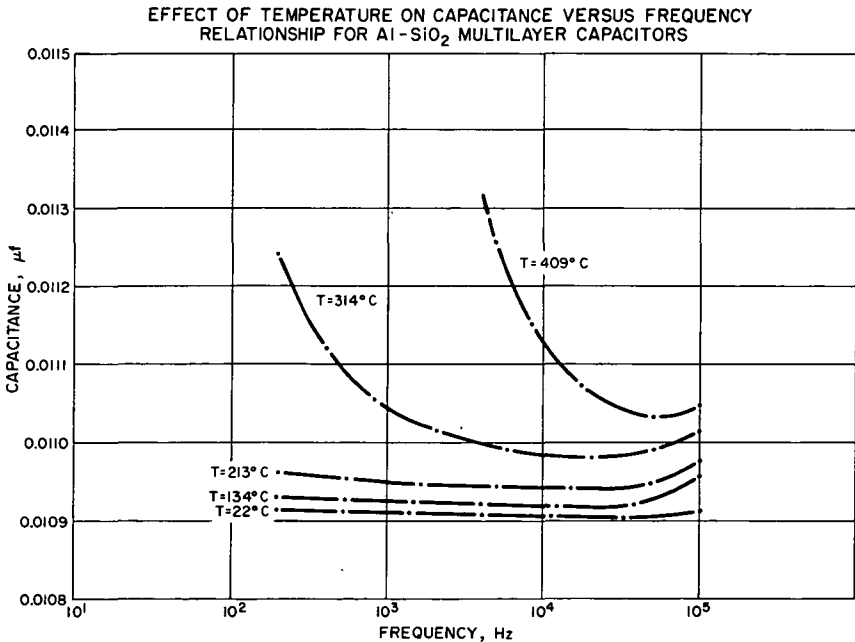


FIGURE 27.—Effect of temperature on capacitance versus frequency relationship for Al-SiO_2 multilayer capacitors.

been greatly used for ac circuits within power processors is the lack of sufficient information on the suitability of currently available capacitors for this. The designer has to turn to specialized manufacturers who can provide him with custom-made components. This process requires the designer to be familiar with capacitor characteristics, properties, and often the feasibility of manufacture. The designer often decides against such a tedious and costly procedure and uses inadequate components designed for signal-level use instead.

Analysis of power networks which include components with nonlinear characteristics calls for a wattmeter with a bandwidth from dc to approximately 1 MHz to indicate the nature of dissipative processes within these components associated with abruptly changing voltage and current waveforms. Currently available wattmeters are restricted to frequencies of several kHz at specific power factors and, therefore, are not suited for this purpose. An experimental instrument that, in addition to the common scale indicator, displays the power dissipation within a certain component as a function of time on a conventional oscilloscope is presently under development (ref. 21). Collector voltage and the corresponding power dissipation in a p-n-p transistor are illustrated in figure 28.

STABILITY

Stability analysis of power processing apparatus is often restricted to iterative experimental observations and adjustments to the design near the completion of a development program. Application of damping methods usually leads to a condition that is considered satisfactory although it deprives the system of the needed capability to react against disturbances that appear on the supply line and the effects of sudden load changes or other dis-

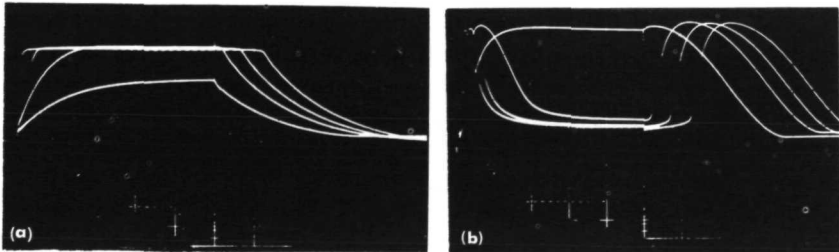


FIGURE 28.—(a) Collector voltage and (b) relative power dissipation of transistor 2N1184A with 2-A collector current. Collector voltage is 5 V/cm, input pulse is 25 μ sec, and base drive is 2, 5, 10, and 20.

turbances that may enter the system. At times, attempts are made to apply the methods of linear feedback analysis to power processors. Unfortunately, the simple and powerful methods of linear control system analysis are not appropriate for networks of power processors, primarily because these methods were developed for systems whose characteristic equations contain real roots that arise from time constants associated with RC and resistance-inductance (RL) combinations within the networks. However, power processing systems include numerous components having nonlinear characteristics near their regions of transition from one state to another. Outside these regions, one could think of these systems as being piecewise linear; however, a piecewise linear system is nonlinear by definition and, therefore, cannot be analyzed by linear methods. This property is underlined by the inherent limitation of a power processor to certain levels of potentials in its input or output stages. The limitation of the power amplifier by its saturation levels is another significant nonlinear characteristic that becomes apparent under conditions of dynamic operation such as turn-on of the power system, abrupt load changes, and other transient disturbances that the power control system should react to and stabilize. A search of the literature for methods of analysis of the aperiodic or nonuniform aperiodic sample data systems soon reveals that no useful method exists that indicates under what specific conditions and by what margin these feedback systems will be stable. Thus, technical information is sorely needed on the dynamic behavior of these feedback systems under steady-state and transient conditions. The solutions to these problems are not simple. When formulated, they lead to differential equations with time-varying coefficients. To complicate matters, the variation of each of these coefficients is determined by the outcome of previous operations.

Intensive work is needed in the field of stability analysis to establish, and provide the designer of power processing apparatus with, the tools needed to ascertain a priori whether a stable system can be realized for certain given conditions. The importance of this work becomes obvious when we realize that it should greatly improve the reliability of the systems and reduce the number of trial-and-error methods currently in use.

A digital computer was used in an attempt to provide the kind of information needed to analyze the stability of a system under given conditions of operation. A power pulse modulator was modeled on a digital computer after its significant parameters were normalized. Data concerning the response to unit step inputs

under practical conditions were obtained and organized to allow the designer to anticipate the degree of stability of the system under given conditions of operation. Figure 29 is the kind of

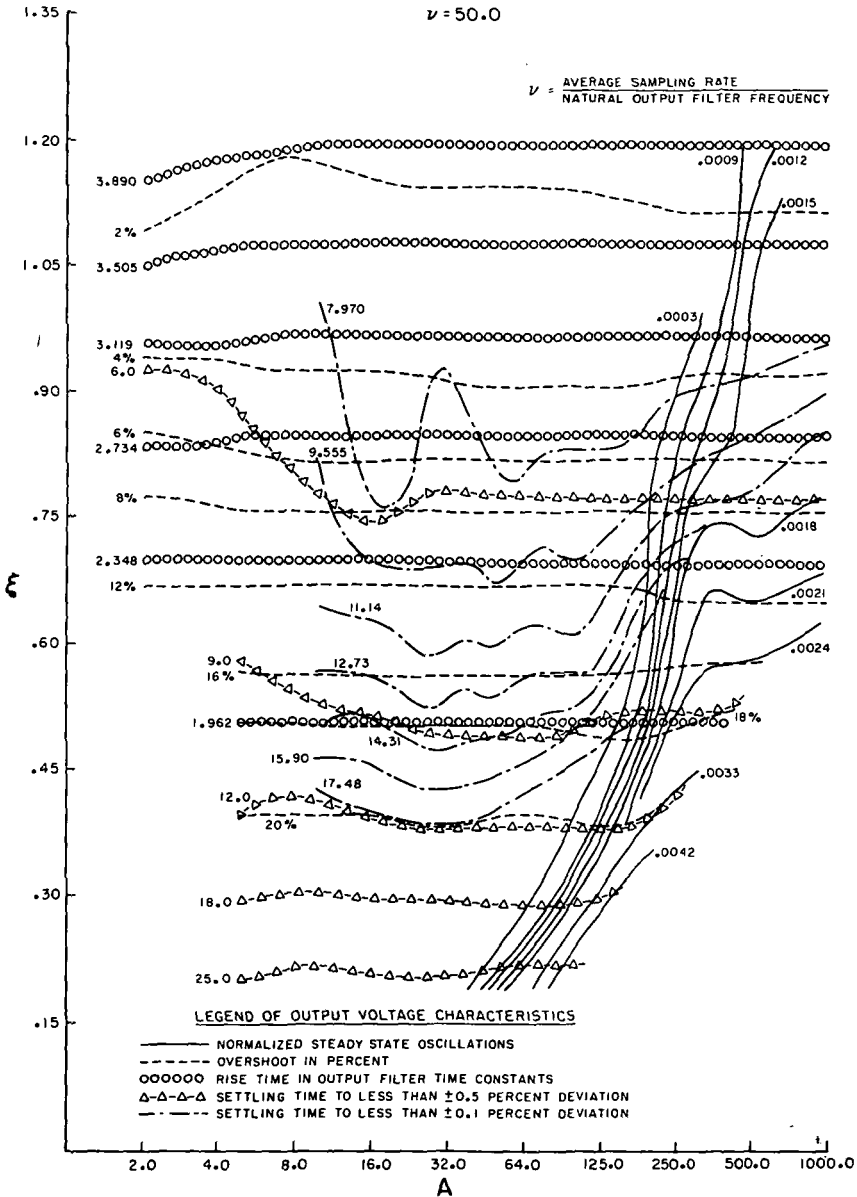


FIGURE 29.—Response of pulse modulator illustrated in figure 9 to unit step input.

graph that can be obtained by this procedure and contains information, for the pulse modulator being considered, on (1) steady-state oscillations; (2) rise time, overshoot, and settling time as functions of the feedback gain; and (3) the damping ratio ξ of an L-type inductance-capacitance (LC) filter; and (4) the ratio ν of average pulse repetition rate to the natural frequency of that filter. Integration of the use of graphs like these into the design process should permit the designer to establish beforehand the stability of his pulse-modulated dc-to-dc converter. A design process based on the significant results of computer modeling should provide detailed information on the significant dynamic properties of these converters and could be part of a deterministic computerized synthesis method rather than to resort to customary methods of computerized network simulation.

EDUCATION AND TRAINING OF PERSONNEL

Engineers in the field of electric power processing come from various areas of electrical engineering. One relatively small group is that of classical electrical power engineers who, at some time in their careers, needed to apply modern static power conversion to the programs within their areas of responsibility. Another smaller group from the area of magnetic amplifiers found better ways to implement the nonlinear philosophies of operation of magnetic amplifiers by way of faster and more flexible semiconductor components. Another small group was supplied by the discipline of control systems engineering. The largest group, however, is from the area of electronic circuit design.

Many electronic circuit designers are at one time or another exposed to the need to stabilize power for sensitive networks. Electronic power networks differ from their signal-level counterparts in the size of voltages, currents, and power dissipation in their individual components. The philosophy of operation of power pulse modulators is closely related to that for pulse modulation in the area of communications technology. The significant difficulties that arise for the electronic circuit designer when entering the field of power processing are rooted in the requirement for efficient operation of power networks. The variable resistive component, the transistor that implements the wave-shaping process, is eliminated. Further, inductors and transformers are introduced for operation at relatively high frequencies; these components are rarely used in conventional signal-level electronics. The networks under consideration consist of inductive, capacitive, and ON-OFF switching elements. No didactic literature

appears available that concerns itself with the quantitative treatment of lossless voltage waveform transformation at power levels and its implementation by nonlinear LC networks under conditions of abrupt switch modulation. The same holds true for textbooks or other teaching material on the stability of pulse-modulated LC feedback networks.

The requirements for competence of electrical engineers in this field are rigorous. They should have a working knowledge of the theory and design of nonlinear active filters that eliminate resistive components for any functional purpose and an ability to analyze these systems under open- and closed-loop operation. These two requirements are imposed by the necessity of producing equipment that will operate without fail. However, no course work concerned with practical attainment of these objectives seems available at any university.

A thorough study of system and control theory to acquire a working knowledge of the advanced tools of mathematics for the analysis of time-varying systems should form a point of departure into the unknown and problematic areas associated with power processing technology. Beyond this, the engineer has to rely on his intuition, which is based on his general training and related experience, to produce working equipment. He must also be able to design pulse, digital, and switching circuits under the aggravating conditions of operation from unstabilized sources. These circuits have reference nodes, whose potentials vary with respect to each other, yet require positive and unfailing transmission of intelligence within the circuits and over the boundaries of the time-varying reference node potentials. The circuits operate, furthermore, near sizable electromagnetic fields that are caused by the abrupt switching of power networks. In addition, they often must operate under adverse environmental conditions. The needed reliability of this circuitry requires positive relation of each function to the past history of the power network and its associated electronic logic in order to assure reliable continuation of operation under conditions of transient disturbance. This eliminates the use of any capacitive coupling between networks and requires the application of complex techniques for the transmission of signals over the barrier of time-varying node potentials. In the area of control electronics, the engineer can expand on his basic education. Nevertheless, extensive training in the general philosophy and physical implementation of analog and digital logic is required in addition to the knowledge that is contained in representative textbooks (ref. 22).

The classical power engineer seems to be more knowledgeable about power components, especially transformers and inductors, than his electronic colleague. Training in this area, however, is unfortunately almost exclusively restricted to utility power applications, a firmly established technology that has remained relatively static for decades. The result is that a study in depth of the electromagnetic processes, and the associated ferromagnetic phenomena, is not a part of the training of engineering students. The power engineer often lacks needed training in the advanced methods of applied mathematics. Often in engineering curricula, more time is devoted to utility power technology than to the broad principles of electrical engineering. The study of transformers, therefore, is often limited to the methods of frequency domain analysis of utility power transformers than to an understanding of the transformer in a broader sense. The power processing engineer requires an understanding of transformers that is based on a quantitative description of phenomena in the time domain because the waveforms and the modes of operation that he encounters are described in this domain. In the analysis of transformers, applying Fourier series expansions of abruptly changing waveforms is extremely cumbersome and obscures an understanding and insight into the quantitative relationships between the significant phenomena. Thus, both the power engineer and the electronic engineer are not trained well enough when they enter, from different directions, into the field of power processing.

This field suffers from the lack of younger talents who are sufficiently interested and capable to make it their profession. A significant impediment for entrance into this field is the lack of understanding of what it is and the belief that anything associated with the notion of electrical power engineering is unchallenging and, worse yet, unglamorous. The requirements for the abilities of an engineer in this field, as developed throughout this report, are severe and require many years of intensive study and training at an unusually high level of sophistication. The encouragement of students to work toward such a goal will have to come from the faculty members of our universities. This, in turn, requires that the problems of this area be brought to the attention of our teachers of electrical engineering to acquaint them with the sophistication and the challenge contained in these problems.

Summary

Processing of electric power has been presented throughout this report as a discipline that draws on almost every field of electrical engineering, including system and control theory, communications theory, electronic network design, and power component technology.

The cost of power processing equipment, which often equals that of expensive, sophisticated, and unconventional sources of electrical energy, such as solar batteries, is a significant consideration in the choice of electric power systems. Its reliability appears in urgent need of improvements since it is believed that the overwhelming majority of failures in sophisticated electric systems originate in the area of power processing. There exists a misconception that this field, which is at this time just emerging in the wake of solid-state components, is well established. This misconception seems to be the cause of the insufficient concern with this field seen in educational institutions and all ranks of engineering management. The valuable work accomplished in the last decade is due to successful attempts by individuals from other areas of electrical engineering to interpret the problems in the light of their abilities and to educate themselves to meet the challenge of these problems. To achieve this, power engineers have ventured into this area and tried to understand and utilize the characteristics of semiconductor components, whereas electronic engineers have tried to apply and modify the theory of signal-level electronic networks.

Every power converter developed today is a piece of art, created by a group of capable individuals, which cannot be modified because of a lack of understanding of the principles involved and their quantitative interpretation. Other areas of technology, for example, communications and control systems, underwent a similar pattern of development in a not too distant past, when, after initial establishment as useful technology, they were permeated with the rigorous methods of mathematical analysis and thus transformed from an art to a science.

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